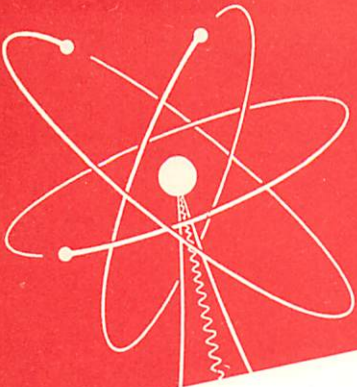

VIDEO I-F AMPLIFIERS

Lesson T2C-5A

DeVRY Technical Institute
4141 W. Belmont Ave., Chicago 41, Illinois
Formerly DeFOREST'S TRAINING, INC.

T2C-5A

VIDEO I-F SECTION

ELECTRONICS TELEVISION
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Chicago 41, Illinois

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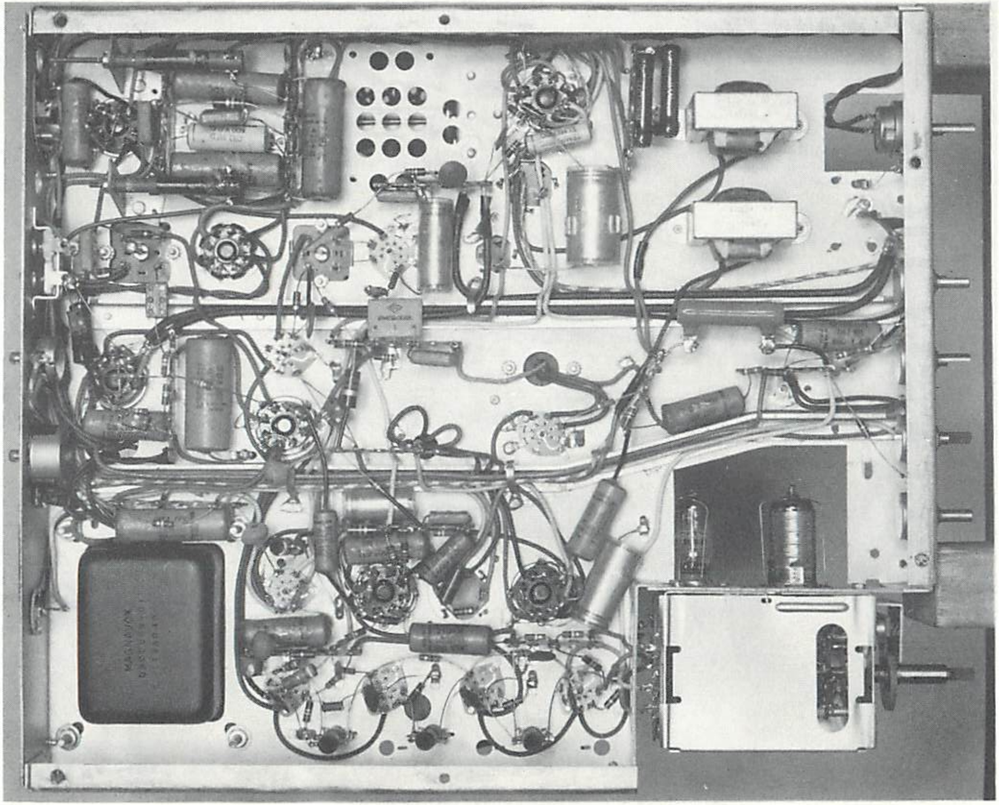

This bottom view of an intercarrier type television chassis shows the i-f section with its miniature tube sockets and stagger-tuned i-f coils immediately to the left of the tuner sub-assembly.

Courtesy Magnavox Company

VIDEO I-F SECTION

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The optimist says his glass is half full; the pessimist says his is half empty.

—St. Louis Post Dispatch

VIDEO I-F SECTION

With the modulated video carrier, modulated sound carrier, and local oscillator inputs; the output circuit of the television tuner mixer stage carries a number of heterodyne frequencies. By means of tuned LC circuits, all except two of these frequencies are attenuated. Known as the "video i-f" and "sound i-f" the two remaining frequencies, together with their sidebands, are increased in magnitude by the i-f amplifier stages.

The most common frequency ranges for these i-f's are the 20 mc and 40 mc ranges. At the present time, it is standard practice to employ video intermediate frequencies of 25.75 mc or 45.75 mc with the sound i-f 4.5 mc lower, either 21.25 mc or 41.25 mc. In any receiver, the lower the intermediate frequency employed, the closer the local oscillator frequency is to that of the r-f tuning circuits. If these circuits couple the oscillator output through the r-f stage into the antenna circuit, its frequency will be radiated and cause interference in other receivers.

It is to reduce this radiation, some receivers employ intermediate frequencies in the 40 to 45 mc range, thereby increasing the separation between the oscillator and r-f circuit resonant frequencies. On the other hand, the lower i-f values provide the advantages of higher amplifier stage gain and stability. As described in detail in this lesson, the video i-f channel includes three to five amplifier stages and the video detector. In addition, most receivers contain circuits for the automatic control of gain.

The receiver i-f amplifier serves to select the desired frequency bands and reject all undesired components of the mixer output, as well as amplify the desired signals to the proper level for application to the respective video and audio detectors. Depending upon the type of receiver, the i-f amplifier may consist of a single or a double series of stages, or a combination of both as mentioned in a previous lesson.

Reviewing briefly, in the dual channel type receivers, the video i-f and sound i-f signals are separated at the mixer output before the signals pass through any i-f amplifiers. In other cases, the two i-f signals are both amplified by one or more common stages before being separated for a modified dual channel system. After separation, each signal is amplified by separate picture and sound i-f stages, respectively.

Thus, the picture and sound i-f amplifiers may be entirely separate, or partly common and partly separate, but so long as the video i-f is finally the only signal applied to the video detector, the receiver is classed as a dual-channel type.

In the intercarrier type receivers, the two i-f signals are passed through a common i-f amplifier system, from the output of which both are applied to the video detector. The video detector causes the i-f carriers to heterodyne and produce the difference frequency of 4.5 mc. Since it carries the audio and video modulation, this difference frequency may be thought of as a second intermediate frequency, which is applied to a sound i-f amplifier tuned to a center frequency of 4.5 mc. By proper filter action in the video amplifier, the 4.5 mc signal does not reach the picture tube to cause interference.

In the sound channel the video amplitude modulation is removed from the 4.5 mc carrier by limiter action after which the a-f signal is recovered by the FM detector, amplified to the proper level, and applied to the loudspeaker.

VIDEO I-F RESPONSE

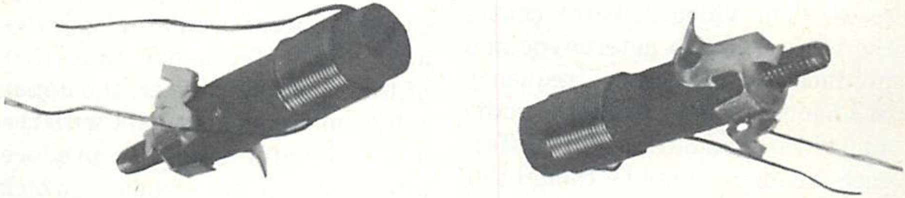
To provide the desired detail in the transmitted picture, video modulation frequencies extending up to 4 mc are employed. In the television receiver mixer output, the video i-f contains sidebands corresponding to those associated with the r-f carrier, and which, therefore, extend to slightly over 4 mc from the i-f carrier. Thus, the video i-f amplifier response must provide for amplification of these sidebands as well as for the i-f carrier.

As you may remember, the received r-f signal consists of the carrier, the upper sideband, and a portion of the lower sideband. Because the local oscillator operates at a higher frequency than that of the incoming carrier, the upper sideband frequencies mix with the local oscillator signal to produce intermediate frequencies which are lower than the i-f carrier, while the lower r-f sideband frequencies result in intermediate frequencies higher than the i-f carrier. Thus, the complete video i-f signal consists of the carrier, a lower sideband extending to approximately 4 mc below the carrier, and a partial upper sideband.

To illustrate this action, in Figure 1, the solid line curve represents the ideal i-f response of a type receiver employing a video i-f of 25.75 mc. Here, the horizontal scale along the bottom represents frequency in mc and the vertical scale at the left represents response in per cent of maximum.

A wide band-pass in the video i-f amplifier provides the advantage of greater detail in the reproduced picture, but tends to lower the signal-to-noise ratio because noise energy is distributed evenly over the frequency band. On the other hand, a reduced band-pass results in less fine detail, but a higher signal-to-noise ratio and greater gain per stage are obtained. In consideration of these factors, commercial monochrome television receivers generally are designed to have a band-pass of anywhere from 3 to 4 mc as indicated by the dashed line and dotted line curves.

In the television receiver, most of the video signal amplification is provided by the i-f amplifier, the gain of which may be anywhere from 1000 to 10,000 or more. Like the r-f section requirements, the tube and circuit capacitances must be kept as small as possible since they tend to reduce stage gain by lowering the L/C ratio.



I-f coils designed for simple impedance coupling in stagger-tuned circuits. The metal clips are used to mount the coil in chassis holes and to hold the powdered iron cores in position. These cores are screwed up or down from the top side of the chassis until the proper inductances are obtained.

D.T.I. Photo

The tuned circuit inductors of all the i-f stages are made variable to permit adjustment of the resonant frequency and generally are overcoupled or resistance loaded to obtain the desired bandwidth. The resulting low impedance of the tuned circuits accounts for the relatively low gain per stage of approximately 20 or less.

Vestigial Compensation

As explained earlier, to provide the desired detail in the transmitted television picture, modulation frequencies extending up to approximately 4 mc are employed. For double sideband transmission, this would require about an 8 mc band for the carrier and both sidebands. However, each allocated channel is only 6 mc wide and, in addition to the modulated video carrier, must include the modulated carrier for the accompanying sound. To meet this condition, the major portion of the lower video sideband is suppressed in the transmitter antenna system by means of a filter designed to completely attenuate all frequencies which are more than 1.25 mc below the carrier.

This type of transmission was referred to in an earlier lesson as vestigial sideband transmission.

Unless the receiver circuits compensate for this condition, the video detector output will be twice as great for video signals below .75 mc as for the higher video frequencies, and amplitude or intensity distortion of the reproduced picture will result.

It will be recalled that the lower transmitted sideband of 1.25 mc is not attenuated at all until, at a point, .75 mc below the carrier, thus the reason for the statement about the resulting video signals being twice as great from 0 to .75 mc or 750 kc as compared to the higher frequency signals, if compensation in the i-f stage is not used.

For example, if the receiver i-f frequency response were made like Figure 2A, which is identical to the transmitted television r-f signal, the signal output from the video detector would have an improper amplitude proportion like that shown in Figure 2B.

It will be noticed that, if the full strength, double sidebands were allowed to exist around the i-f carrier, those video signals from 0 to 1.25 mc. would be greater in amplitude than the higher picture frequencies. In fact, as shown in Figure 2B, those frequencies which get full double sideband amplification from 0 to 750 kc are twice as great in amplitude as for those above 1.25 mc where only the single sideband exists.

To show the additive result of the lower-frequency double sidebands from an incorrect i-f setting such as in Figure 2A, the diagram has been divided into percentages of maximum amplitude, with the sideband frequency distance away from the i-f carrier noted in mc.

As the 1.25 mc sideband points in Figure 2A, 0% of the lower sideband plus 100% of the upper sideband provides a signal output equivalent to the 100% point as shown for a picture frequency of 1.25 mc in Figure 2B. Continuing this comparison, at the 1 mc sideband points in 2A, approximately 50% of the lower plus 100% of the upper, provides a signal output amplitude of 150% at 1 mc in 2B. At the .75 mc sideband points in 2A, 100% + 100% provides an output amplitude of 200% at .75 mc in 2B. The same holds true for .5 mc, .25 mc and on down to the unmodulated or zero reference. Thus, as was stated, all the low video frequencies below .75 mc would receive twice as much amplitude as the higher single side band video signal components would obtain under these conditions.

As shown in Figure 3A, to equalize this possible vestigial sideband effect, the receiver video i-f tuned circuits are adjusted so that, at the exact video i-f carrier frequency, the gain or response is down to 50% of that for the single sideband frequencies which are above 1.25 mc that represent the higher video signal components. With this i-f alignment compensation in the response curve of Figure 3A, the detected output is the curve of Figure 3B.

For example, the added amplitude percentages of the double sideband from 3A producing Figure 3B are as follows: at the 1.25 mc sideband points of Figure 3A, 0% of the lower sideband plus 100% of the upper sideband provides the 100% detected output as shown for a picture frequency of 1.25 mc in Figure 3B. At the 1 mc sideband points in 3A, approximately 10% of the lower plus 90% of the upper sideband provides a signal output amplitude of 100%.

At the .75 mc sideband points in 3A, 20% of the lower plus 80% of the upper again makes a 100% total amplitude in 3B. Continuing for the remaining lower sideband frequencies, at .5 mc, 30% + 70% equals 100% and at .25 mc, 40% + 60% equals the same 100% result in Figure 3B.

Thus, the compensation i-f alignment response curve of Figure 3A provides for a flat detected output from 0 cycles on up at the output of the video detector, as shown in Figure 3B. This condition affords excellent picture reproduction since all original low, medium, and high picture frequencies are reproduced in proper magnitudes.

If the oscillator fine tuning control in the tuner is not adjusted properly, it is possible that the resulting video i-f out of the mixer might be shifted upwards to the right in Figure 3A causing the lower frequencies to be accented. This would produce an intense contrast on the larger or low frequency components of the picture, with inadequate comparison of the high frequency components present. The result would be a more intense picture with less detail.

For example, if the mixer i-f output were shifted .5 mc to the right on the curve of Figure 3A by misadjustment of the oscillator fine tuning control, it would have a starting or zero reference on the 70% point of the curve, rather than 50%. A ±.25 mc sideband distance from this point would add 80% and 60% for a 140%

total amplitude, which at this comparatively low frequency of .25 mc, or 250 kc, is 40% over the desired flat response in Figure 3B. The results for the other lower frequency sidebands would thus prove the same.

If the tuner fine tuning control were misadjusted in the opposite direction, then the low frequency response of the picture would be reduced as a result of the video i-f shifting down to the left on the curve of Figure 3A. This would cause poor contrast of the larger components of the picture.

For example, a shift to the left of .5 mc on the curve of Figure 3A would place the i-f starting or zero reference on the 30% point of the curve, instead of the desired 50% point. A ±.25 mc sideband distance from this 30% point would add 40% + 20% for a 60% total amplitude, which, at this low frequency of 250 kc, is 40% below the desired response in Figure 3B. This would result in poor picture synchronization and generally poor low-frequency component response.

Since the tuner fine tuning control can shift the resulting i-f's at the mixer output from a range of 2 mc and up, it is necessary to know the importance of this control in its effect on the i-f stages and that it should be fine adjusted with the detail balance of the picture in view.

Video I-F Selectivity

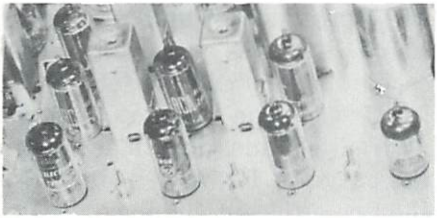
Because of the broad bandwidth of the tuner, it is possible that channel frequencies which are next to, or adjacent to the station tuned in may appear as interference to be detected along with the normal i-f carriers of the desired channel.

For example, Figure 4 shows the relationship of a selected station, such as channel 3, to its adjacent carrier frequencies for channels 2 and 4. Assuming a 25.75 mc video i-f and a 21.25 mc sound i-f, this receiver's local oscillator frequency for channel 3 is 87 mc. The carriers for the desired channel are marked on the curve of Figure 4 along with the two adjacent channel carriers which could possibly interfere.

Although the FCC does not assign adjacent channels to television stations operating in the same area, certain fringe-area receiver locations may be troubled with this condition of overlapping signals from adjacent carriers caused by transmitters operating on adjacent channels in different areas.

Notice in the diagram of Figure 4 that the adjacent carriers present, which are most likely to cause interference, are the ones closest to the selected channel frequencies. That is, the sound carrier on the channel below and the video carrier on the channel above are most likely to cause trouble since they appear within the indicated tuner response curve represented by the dashed line response curve in Figure 4.

The other adjacent channel carriers, not shown in the diagram, will fall outside the tuner response curve and thus, cannot get through to heterodyne with the existing local oscillator frequency.



The i-f strips of a typical television receiver. Notice the clips and core screws for the stagger-tuned video i-f coils in the foreground between the tubes.

Courtesy Wilcox-Gay Corp.

Thus, in this case, when the tuner is turned to channel 3, any channel 2 adjacent sound carrier signal received will have a frequency of 59.75 mc and, in heterodyning with 87 mc, will produce an adjacent i-f difference of 87 mc − 59.75 mc, or 27.25 mc as noted in Figure 4; while any adjacent channel 4 video carrier signal received at 67.25 mc, will beat with the 87 mc local oscillator frequency and produce an adjacent i-f difference, as noted in the diagram of 87 mc − 67.25 mc, or 19.75 mc. The desired video i-f and sound i-f are also indicated.

Of necessity then, the television receiver i-f section should be selective enough in its tuned circuits to be able to discriminate against any of these undesirable adjacent intermediate frequencies that are referred to in Figure 4.

It is recalled from a previous lesson, that the intercarrier type of receiver i-f system is broad enough to amplify both the sound and video i-f carriers of a particular channel into the video detector circuit where the 4.5 mc sound i-f is developed. The response curve of this intercarrier system appears in Figure 5A and, in this case, shows the result of using tuned circuits to exclude or trap out the unwanted adjacent i-f's noted on the diagram of Figure 4.

These adjacent i-f's of 19.75 mc and 27.25 mc are located at the minimum points on the i-f response curve, while the desired sound i-f, 21.25 mc, is at a less than 10% point and the desired video i-f is at the previously mentioned 50% point.

By comparison, the tuned circuit response curve for the dual channel type of receiver i-f system is shown in Figure 5B. The one definite difference lies in the fact that the sound i-f of the desired channel also is reduced as much as possible in the video i-f output.

It will be recalled that, in the description of the dual channel type of i-f system, the sound i-f signal is removed from the video i-f stages before it appears at the video detector. From the point of removal the sound signal is directed into its own channel. Little, if any, sound i-f signal is allowed to remain in the latter video i-f stages.

Thus, a large part of the television receiver selectivity for either type of i-f system is furnished by the video i-f stages. The tuned circuits must be designed to provide a bandwidth of approximately 4 mc and, at the same time, have sufficient attenuation just beyond the bandwidth limits so that the adjacent sound and video carriers do not cause interference at the picture tube. In addition, the desired sound i-f carrier must also be attenuated to a designated point depending on the type of i-f system in the receiver.

Wave Traps

To attenuate or reject these various i-f signals the desired amount, many television receivers employ special series or parallel tuned circuits associated with the i-f amplifier interstage coupling or tube cathode circuits, known as **wave traps.** Each of these sharply tuned LC circuits either rejects, absorbs or provides a low impedance path to ground for the particular signal to which it is tuned, such as the adjacent sound i-f, or adjacent video i-f, or the tuned channel sound i-f.

Such traps are not only used for the attenuation of these i-f signals, but also are employed to reduce the "used" channel sound i-f in the intercarrier system to less than 5% of maximum response, and occasionally, to place the used channel video i-f at the 50% response level as explained for Figure 3.

INTERSTAGE COUPLING CIRCUITS

The i-f amplifier circuits must be designed to satisfy these requirements of band-pass, gain and selectivity. Usually, these circuits are designed either in terms of band-pass filter theory, or a system of stagger-tuned single coil i-f stages. In both cases, a combination of both methods is used.

Some of the following portions of this lesson make references to certain design characteristics of i-f circuits. These references provide a general idea and appreciation of what is involved in the specifications of some of these various circuits.

Band-Pass Filters

A simple band-pass filter consists of a number of tuned circuits and a terminating resistor. The L and C of the tuned circuits are determined by the resistance of the terminating resistor, the frequency limits of the passband and some particular frequency outside the band limits at which great attenuation is desired. In this type of filter, there is no coupling of any kind between the various components.

The type of band-pass filter employed in video i-f circuits is known as a coupled filter, and consists of two tuned circuits which are coupled in such a way that the circuit has maximum response at two frequencies. This arrangement is illustrated in Figure 6A which shows a double-tuned coupling transformer, L_pL_s, both primary and secondary of which are tuned to the same frequency and closely coupled to provide the desired double-peaked response and wide bandwidth such as indicated by the response curves of Figure 5.

Generally speaking, coupling may be defined as a means of transferring energy from one circuit to another, and any coupled filter contains a component called the mutual coupling impedance which is common to both circuits and by which the energy transfer is made. Depending upon the coupled filter arrangement, THE MUTUAL COUPLING IMPEDANCE MAY BE EITHER INDUCTIVE, CAPACITIVE OR RESISTIVE.

In the circuit of Figure 6A, the magnetic lines of force, produced by the primary current, couple L_p and L_s and may be represented as the coupling inductance, L_m. Thus, L_m is the mutual coupling impedance and its value in henrys is determined by the extent to which the lines of force cut the secondary. Theoretically, when all the flux set up by L_p cuts the turns of L_s, then L_m would equal the square root of the product of L_p and L_s. Written as an equation:

$$L_m = \sqrt{L_pL_s},$$

in which:

L_m = mutual inductance in henrys
L_p = primary inductance in henrys
L_s = secondary inductance in henrys.

However, this situation is impossible to obtain in practice and the actual flux which cuts the secondary is equal to some fraction, k, of the total flux. Thus, for any practical coupled filter, the mutual coupling inductance is:

$$L_m = k\sqrt{L_pL_s},$$

in which L_m, L_p, and L_s are as in the preceding equation and

k = some number between 0 and 1.

Called the coefficient of coupling, k depends upon the spacing between L_p and L_s, the relative positions of these coils, and their lengths and diameters. When this last equation is rearranged:

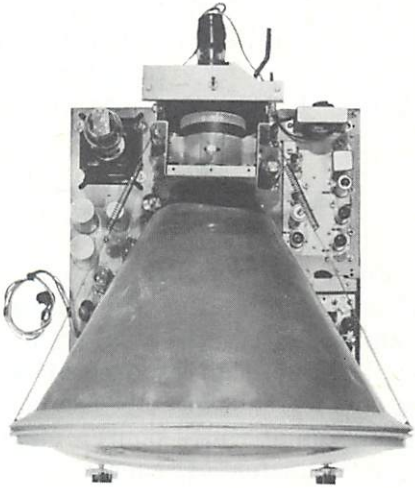
$$k = \frac{L_m}{\sqrt{L_p L_s}}$$

It shows k to be the ratio of the mutual inductance actually present (L_m) to the maximum value theoretically obtainable, ($\sqrt{L_p L_s}$). For example, when the primary and secondary coil spacing is reduced, L_m is increased and the coefficient of coupling is greater.

Mutual inductance is a difficult quantity to control in assembly line production of magnetically coupled transformers like that of Figure 6A, hence in some receiver designs, r-f tuners and i-f amplifier coupling circuits employ an arrangement similar to that shown in Figure 6B. Here, an actual coil, L_m, provides the mutual coupling impedance and, although different in appearance, this coupling network is equivalent to the transformer of Figure 6A.

In the circuit of Figure 6B, the primary, L_p and secondary L_s, are arranged so there is little or no magnetic coupling between them. In series, L_p and L_m form the inductive component of the primary tuned circuit, the capacitive component of which is the V_1 output capacitance C_o.

In similar manner, L_s and L_m form the inductive component, and V_2 input capacitance C_I is the capacitive component of the secondary tuned circuit. In the primary circuit, the signal currents produce an emf across L_m, and this small emf causes currents in the secondary circuit which induce larger signal voltages across the complete tuned secondary circuit.



A top view of an intercarrier type television receiver chassis. The video i-f section is located on the outside right of the chassis, just beyond the tuner. The video amplifier and 4.5 mc sound i-f section are the 3 miniature tubes going from back to front adjacent to the other section.

Courtesy Motorola, Inc.

Capacitor C_b is inserted in series with L_s to prevent the V_1

plate voltage from being applied to the V_2 grid, while C_s provides a signal path to ground from the B+ end of L_m. At the intermediate frequencies, the reactance of C_b and C_s are so low that their effects on the circuit tuning are negligible.

As the inductance of L_m is added to both circuits, the coefficient of coupling for the arrangement of Figure 6B may be expressed as:

$$k = \frac{L_m}{\sqrt{(L_p + L_m)\ (L_s + L_m)}}$$

For coupling circuits like those of Figure 6, as the coefficient of coupling k is increased from a small value, the response at the resonant frequency, f_r, rises to a maximum at which point a maximum transfer of energy takes place between the primary and secondary. At this time, the coefficient of coupling is called the critical coupling, k_c, of the circuit.

For example, with a further increase of coupling in a greater L_m, a second peak appears, thus forming a double humped curve. With continued increase in k, these peaks of maximum response become more pronounced and move farther and farther apart.

Since k determines the frequency difference, or spacing between the two maximum response peaks, it can be chosen to provide the desired wide passband for the video i-f amplifier coupling circuit. The direct relationship between the passband and coefficient of coupling is shown by the following equation:

$$\Delta f = kf_o$$

in which:

Δf = the width of the passband, and is read as "delta f."

f_o = the center frequency of the passband.

Therefore, L_m is chosen to give the desired coupling coefficient, k, while L_p and L_s are selected to resonate the two tuned circuits to some frequency, f_r, near the upper limit of the i-f band. Thus, with a relatively large k, the response curve has one peak above the center frequency, f_o, and a second peak at some frequency spaced the same distance below f_o.

For any particular transformer arrangement like those of Figure 6, the response curve shape depends upon the ratio of the actual degree of coupling, k, to the critical value, k_c. This ratio, k/k_c, is called the "shape factor" and, in video i-f coupling circuits, it is common practice to employ values for k which range from 1.5 to 2.0 times the critical value.

Since the required k is determined by the desired passband and center frequency as explained above, the only way in which the shape factor can be changed is by varying the critical coupling, k_c. For any particular set of coils, k_c is determined by the respective coil Q's, and the relationship may be expressed by the following equation:

$$k_c = \frac{1}{\sqrt{Q_p Q_s}}$$

in which:

Q_p = the Q of the primary coil

Q_s = the Q of the secondary coil.

As the Q depends upon the resistance associated with a tuned circuit, loading resistors such as R, in Figure 6A, and R_1 and R_2, in Figure 6B, are employed to reduce the circuit Q's to the degree necessary to provide the desired k_c. In some designs, only the secondary is resistance loaded, while in others, both the primary and secondary are loaded.

For a satisfactory shape factor, the approximate resistance of loading resistor R can be obtained from the following equation:

$$R = \frac{1}{2\pi \Delta_f C_t},$$

in which:

Δf = the width of the passband

C_t = the total capacitance in shunt with the coil to be loaded.

The amplifier gain is directly proportional to the resistance of R, and therefore it is inversely proportional to Δf and C_t. That is, the wider the passband and the greater the circuit shunt capacitance, the lower the i-f stage gain.

Transformer Coupled Circuit

The schematic diagram of the typical i-f amplifier employed in one well known television receiver is shown in Figure 7. In this circuit, mixer tube V_1; i-f amplifier tubes V_2, V_3, V_4, and V_5; and video detector tube V_6 are coupled by means of over-coupled transformers, each of which has a resistor connected across its secondary to provide the desired band-pass characteristics.

With the exception of L_1, each transformer coil is resonated to the desired frequency by the combined capacitance of the tube and a small trimmer capacitor, which provides for slight adjustments of the resonant frequency when aligning the circuit. In many i-f amplifiers, all the capacitances are fixed and the circuits are tuned by permeability adjustment of the coils.

In Figure 7, the V_2, V_3 and V_4 grid circuits are completed to ground through capacitors C_{18}, C_{24} and C_{29}. A variable negative d-c voltage can thus be applied to the V_2, V_3 and V_4 grid circuits from the slider on contrast control potentiometer P for contrast control.

The contrast control of a television receiver determines the strength of the video signal to be applied to the picture tube input. It controls the contrast between the light and dark areas of the reproduced image. The contrast control is operated manually and may be inserted in either the video amplifier or the video i-f amplifier. When used in the i-f section, as it is here, the contrast control is employed to vary the grid bias on most of the i-f amplifier tubes and, thereby, vary their operating points and stage gain.

The required d-c voltage across the contrast control often is obtained from a negative d-c source in the power supply, although some other circuit may supply it. Connected across this voltage source in Figure 7, the voltage divider R_{16}, P, and R_{20} provides one fixed and one variable negative d-c voltage for bias. V_2, V_3, and V_4 grid circuits are returned through resistors R_4, R_{11}, and R_{18} to the slider of potentiometer P, thus allowing their bias voltage to be varied so that the desired contrast is obtained in the reproduced image.

The changing of the i-f amplifier tube grid bias causes a slight variation in the tube input capacitance and therefore, some detuning of the associated tuned circuit. To counteract this change of input capacitance, cathode resistors R_5, R_{10} and R_{19} are unbypassed to provide a small amount of degeneration.

To provide the desired rejection of the adjacent channel and accompanying sound intermediate frequencies and to attenuate the video intermediate frequency to the desired 50% amplitude, several trap circuits are included in this amplifier.

In the V_1 plate circuit, a trap consisting of coil L_3 and capacitors C_3 and C_{16} is connected between the junction of L_1 and L_{16} and the junction of L_{16} and R_1, this latter point being bypassed to ground by C_{17}. Inductively coupled to L_1 and tuned to the adjacent sound i-f, the trap absorbs energy at this frequency, and as can be seen from Figure 5, is close to the video i-f thereby assisting in reducing the video i-f response to the 50% point on the over-all response curve as explained previously. Because it absorbs energy from L_1, it is known as an absorption trap and its effect is shown at the 27.25 mc point in Figure 5.

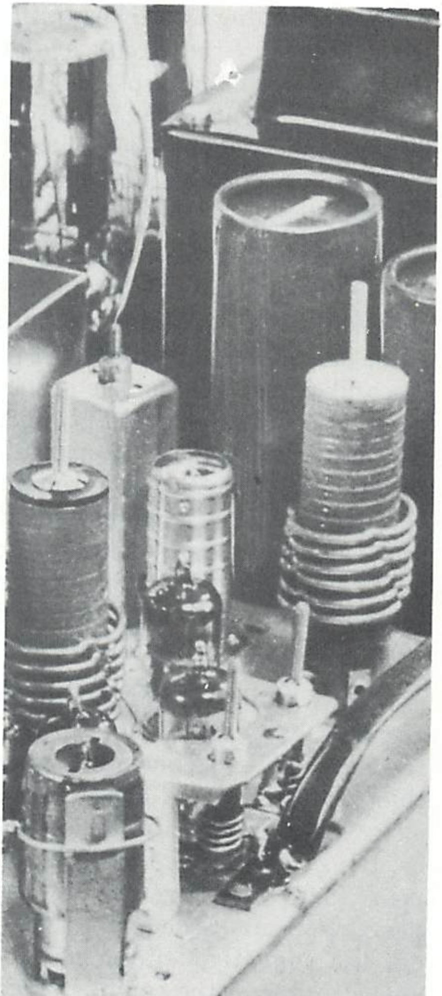
Magnetically coupled to the primary of the second i-f transformer, coil L_6 with capacitors C_6 and C_{21} forms a trap circuit which is resonant to the adjacent video i-f. $L_6C_6C_{21}$ also is known as an absorption trap since, at the frequency to which it is tuned shown as 19.75 mc in Figure 5, it absorbs energy from the coupling circuit.

The desired sound i-f signal is removed from this video i-f channel and applied to the sound channel by means of absorption trap $L_9C_9C_{27}$, which is inductively coupled to the primary winding L_7 of the third video i-f transformer and tuned to absorb the channel sound i-f for its own circuits.

One end of this trap is connected directly to the control grid of the first sound i-f amplifier tube, thereby providing a directed outlet for the desired sound i-f. The other end of the trap is returned through R_{15} to the junction of R_{16} and P_1. Thus, the fixed negative voltage at this junction provides the sound i-f amplifier tube with bias. The sound signal a-c path to ground is completed through a filter capacitor C_{28}, which is connected between ground and the junction of trap coil L_9 and isolation filter resistor R_{15}.

Due to its coupling to L_7, the trap absorbs sound i-f energy from the video i-f section to help produce the characteristic dual channel i-f response curve of Figure 5B.

Specifically, Figure 7 is a modified dual channel arrangement since V_2 and V_3 are common to both video and sound i-f's, with the sound i-f take-off trap appearing in the output of the V_3 stage. R_{14}, in the trap circuit, is



In order to have high Q coils, the sound traps in this receiver use a heavy wire, wound on a large coil form. These traps are also core tuned.

Courtesy United Scientific Labs.

employed to load the trap and thus broaden its bandwidth so that it will pass the FM sound band on to the receiver sound channel.

The remaining two traps in the latter stages of the circuit are of the series resonant type and are also tuned to the sound i-f in order to eliminate any possibility of its interfering with the picture information.

The first such trap consists of L_{12}, C_{12} and C_{32} and is connected between the V_5 grid and ground, while the second consists of L_{15}, C_{15} and C_{37} and is connected between the V_6 cathode and ground. Although not readily apparent, these traps are series resonant circuits, and the combination of two capacitors and one coil for each trap provides much sharper tuning than could be obtained with a simple LC circuit.

In both traps, the parallel tuned circuits, $L_{12}C_{12}$ and $L_{15}C_{37}$, are resonant at a frequency higher than the sound i-f. Thus, to the exact sound intermediate frequency, the parallel tuned circuits appear inductive. By varying the parallel circuit resonant frequency, the inductive effect may be varied to bring the complete trap circuit into series resonance with the series capacitors, C_{32} or C_{15}. Since the trap impedance is minimum at the resonant frequency, it offers a ready path to ground for any sound i-f that might still be present.

Mutual Impedance Coupled Circuit

To illustrate the use of both coupled filters and capacitive coupling Figure 8 shows the i-f amplifier employed in another nationally advertised television receiver. V_1 is the mixer; V_2, V_3, and V_4 are i-f amplifiers; and V_5 is the video detector. Coupled filters similar to that of Figure 6B are employed between V_1 and V_2, and V_2 and V_3, while capacitive coupling is employed between V_3 and V_4, and V_4 and V_5.

In the coupled filter between V_1 and V_2, L_1 and L_3 in series, resonate at the desired frequency with the V_1 output capacitance, L_2 and L_3 in series, resonate with the V_2 input capacitance, and L_3 provides the mutual coupling impedance. Coils L_1 and L_2 have movable, powdered iron cores that permit the adjustment of the resonant frequency to the desired value. The d-c blocking capacitor C_1 prevents the V_1 d-c plate voltage from being applied to the V_2 control grid while allowing the signal to pass freely.

Resistors R_1 and R_2 load the tuned circuits to widen the coupling circuit response, while bypass capacitors C_7 and C_8 return the lower ends of L_3 and R_2 to ground. A variable, negative d-c voltage from an automatic gain control circuit, the operation of which is described later, is applied in series with R_{10} and R_2 to the V_2 control grid.

In like manner, L_4 and L_6 resonate with the V_2 output capacitance, L_5 and L_6 with the V_3 input capacitance, and L_6 provides mutual coupling between the primary and secondary circuits. C_2 prevents the V_2 d-c plate voltage from being applied to the grid of V_3, while R_4 and R_5 broaden the response. Signal paths to ground are provided by C_{10} and C_{12}, and as indicated, an agc voltage is supplied in series with R_5 and R_{11} to the V_3 grid circuit.

The circuit between V_3 and V_4 employs capacitive coupling by means of C_3, the capacitance of which is large enough to make its reactance very low at the intermediate frequency. Coils L_7 and L_8 are resonated at the desired frequencies with their respective tube capacitances and are loaded by resistors R_7 and R_8. There is no inductive coupling between them. In the V_3 plate circuit, the signal path from L_7 to ground is completed by bypass capacitor C_{13}.

In series with C_3 between V_3 and V_4, traps L_9C_5 and $L_{10}C_6$ are tuned respectively to the adjacent sound i-f and the adjacent video i-f. Designed so that their Q's are very high, these traps are very sharply tuned and present a very high impedance to both adjacent i-f signals, but offer very little impedance to the video i-f signal. Thus, very little, if any, interfering i-f signals pass through the coupling circuit, while the video i-f and sound i-f signals pass through properly, for an intercarrier i-f system.

The V_4 output is coupled through C_4 to the input circuit of video detector V_5. Resonated with their respective shunt capacitances, coils L_{11} and L_{12} form the two sections of a coupled filter in which the degree of coupling may be adjusted by varying the capacitance of C_4.

The bandpass filter method of interstage coupling employed in Figures 7 and 8 provides higher gain per given bandwidth because the output capacitance of one tube is isolated from the input capacitance of the following one. Hence the various tuned circuits can have relatively high inductances.

Stagger Tuning

Because of its simplicity and ease of adjustment, another method of interstage coupling, known as **stagger-tuning,** has come into common use in wideband i-f amplifiers. In this method, a single tuned circuit acts as the common coupling impedance between each pair of tubes and produces a single peaked response curve. In a television receiver, several single tuned stages are employed, with each adjusted to a different frequency within the desired bandwidth so that, when all the individual response curves are combined, the result closely resembles the proper response curve.

Figure 9A shows the diagram of a two stage impedance-coupled, stagger-tuned i-f amplifier, in which L_1 and C_1 form a tuned circuit that is parallel resonant for maximum load impedance at one frequency within the i-f passband, and L_2 and C_2 constitute the second tuned circuit which is parallel resonant for maximum load impedance at a different frequency.

The transformer-coupled dual channel television receiver i-f sections. The transformers are mounted inside the shield cans between the i-f tubes.

Courtesy Du Mont Laboratories

C_1 represents the sum of the V_1 output capacitance, the V_2 input capacitance and the circuit stray capacitance. Likewise, C_2 represents the sum of the V_2 output, the V_3 input and the circuit stray capacitances. Coils L_1 and L_2 contain movable, powdered iron cores that permit varying the coil inductance so that resonance may be adjusted to the desired frequency.

The i-f signal is coupled between tubes by means of capacitors C_3 and C_4, which have very low reactance at the intermediate frequency. Bypass capacitors C_5 and C_6 complete the signal paths to ground from the B+ ends of L_1 and L_2. The reactances of C_3, C_4, C_5, and C_6 are so low that resistors R_1 and R_2, in addition to completing the grid circuits to ground, are effectively in parallel with their associated tuned circuits, and thereby reduce the circuit Q and broaden the response curves.

In the design of an individual tuned circuit, bandwidth is assumed to be the frequency difference between two points, equidistant above and below resonance, at which the amplitude is some stated percentage of the resonant frequency amplitude. Generally this is taken as 70.7% of resonance frequency maximum. When designing an i-f amplifier having several stages, the calculations for the over-all bandwidth also are made on the basis of the 70.7% of maximum overall response.

In the case of a single stage tuned amplifier, employing one tuned circuit that is loaded by a low resistance, such as shown in Figure 9A, the bandwidth between the 70.7 percent points is given by:

$$\Delta f = f_r \left(\frac{X_r}{R}\right) = \frac{f_r}{Q_e}$$

in which:

Δ_f = bandwidth

f_r = resonant frequency

X_r = reactance of inductive or capacitive element of the tuned circuit at the resonant frequency.

R = loading resistance

$Q_e = \frac{R}{X_r}$ = effective Q of a heavily loaded tuned circuit.

When the amplifier system has two or more stages, all of which are tuned to the same frequency, the over-all amplification characteristic is the product of the amplification characteristics of the individual stages, and therefore, it is a narrower bandwidth than the individual stages.

For example, if both amplifier stages of Figure 9A were identical and tuned to the same frequency, the total gain would be equal to the square of the gain of one stage. However, the 70.7% points of the individual stage response now reduces to the $(.707)^2$ or 50% points on the two stage amplifier output response. The 70.7% points of the two stage amplifier with identical resonant frequencies, f_r, will be closer to each other and correspond to a bandwidth that is only about 64.4% of the single stage bandwidth. Therefore, in the circuit of Figure 9A, if each stage had a gain of 10, a bandwidth of 4 mc, and both were tuned to the same frequency; the over-all gain would be 10^2 or 100, while the over-all bandwidth would be .644 × 4 or 2.56 mc.

This reduction of bandwidth is undesirable in the television receiver i-f amplifier since it reduces the fine detail in the reproduced image. Thus, as more stages are added to the amplifier, if the desired bandwidth is to be maintained, the Q of the individual stages must be decreased by reducing the resistance of the shunting or loading resistor R. This resistance is the maximum impedance that the tuned circuit can have under any condition and when it is decreased, the amplifier gain per stage, which is given by $A_v = g_m R$, is reduced also, therefore more stages are required to provide the desired output.

As stages are added, the reduction of bandwidth may be overcome by tuning the alternate stages slightly above and below the center frequency of the i-f passband, in a method known as stagger tuning. For example, in a two stage amplifier, if the two circuits are tuned to frequencies that are equidistant above and below the center frequency, f_o, and are separated, such as in Figure 9B from f_A to f_B, by an amount equal to the desired bandwidth, Δf, the over-all bandwidth has been found to be 1.4 times that of the one stage.

As indicated in Figure 9B, because of the off-tuning, the amplification of each stage at the center frequency, f_o, is 70.7% of the amplitude at its resonant frequency value f_A or f_B and the over-all gain is equal to only one-half the square of that of a single stage. In Figure 9A, when each stage has a gain of 10, a bandwidth of 4 mc and is tuned respectively to $f_o + .5\ \Delta f$ and $f_o - .5\ \Delta f$, the overall gain is $.5 \times 10^2$, or 50, and the bandwidth is 1.4×4 or 5.6 mc.

However, to meet the rejection requirements of the television receiver i-f amplifier, the over-all bandwidth must be reduced from 5.6 mc to 4.0 mc, a reduction of 29%. This may be accomplished by diminishing the bandwidth of each tuned circuit by the same amount.

Hence, if the bandwidth of each stage is reduced 29%, that is by a factor of $1\Delta f - .29\Delta f$ or to $.71\Delta f$ of each original stage Δf of 4 mc, the resonant frequencies of the stages then can be equal to $f_o + 1/2\ (.71\Delta f)$ and $f_o - 1/2\ (.71\Delta f)$. This is simplified to $f_o + .355\Delta f$ and $f_o - .355\Delta f$. The over-all bandwidth for the two stages will then be equal to the constant, 1.4, times $.71\Delta f$ which is $.994\Delta f$ and approximately equal to the desired over-all response.

When under the above conditions, an over-all bandwidth, Δf, of 4 mc is desired for 2 stages, then each stage, first of all, should be designed to have an individual bandpass of 71% of 4 mc or 2.84 mc. Then $.355\Delta f$ which determines the frequency distance from f_o, is equal to $.355 \times 4$ or 1.42 mc.

Thus, as shown in Figure 9B, the resonant frequency, f_A, for the first stage will be equal to f_o of 24 mc − 1.42 mc or 22.58 mc. The resonant frequency, f_B, for the second stage will be equal to 24 + 1.42 mc or 25.42 mc.

The over-all passband, curve C of Figure 9B of these two stages, should then be equal to $.994 \times 4$ mc or 3.98 mc, which is the result of having made the response of each stage equal to 71% of 4 mc. The final over-all result of 3.98 mc is indicated by Δf, from approximately 22 mc to 26 mc in Figure 9B.

In this stagger tuning arrangement, the bandwidth of each stage, represented by curves A and B in Figure 9B, has been narrowed but, applied together, still provide proper over-all response. The gain of each stage therefore is greater than that which would exist if it had to pass 4 mc.

In narrowing the response from 4 mc per stage to 2.84 mc, as previously mentioned, the Q of each stage is increased. At these particular values, that resulting increase is equal to a factor of 1.4 times and can be obtained by increasing the resistances of the shunt resistors, R_1 and R_2 in Figure 9A by 1.4 times the resistance they would have when providing the broader 4 mc bandwidth per stage.

This 1.4 times rise factor in Q also causes a like increase in stage gain which is 1.4 times the previously mentioned gain of 10. Thus, from a previous equation, the overall gain of the two stages is equal to $.5 \times (1.4 \times 10)^2$ or 98. This is indicated at the approximate 71% points of curve C in Figure 9B from the over-all voltage gain values at the right of the graph.

In summary then, two stages tuned to the same frequency with a bandpass of 4 mc each, and a gain of 10, produce an over-all result in a gain of $(10)^2$ or 100, but a bandwidth of only 2.56 mc which is too narrow. The same two stages with the same individual bandpass of 4 mc and gain of 10, but stagger tuned to $f_o + 2$ mc and $f_o - 2$ mc, or 4 mc apart will have an over-all gain of $.5\ (10)^2$ or 50 and a total bandwidth of 5.6 mc which is too wide.

Finally, the same two stages again but, this time, each having a somewhat narrower bandwidth by 29% and a resulting increased gain to 1.4×10 or 14, are stagger tuned to $f_o + 1.42$ mc and $f_o - 1.42$ mc, or 2.84 mc apart, will have an over-all gain of $.5\ (14)^2$ or 98 and a total bandwidth of 3.98 mc which is approximately the de sired total bandwidth of 4 mc.

Stagger-Tuned Circuit

In order to produce some desired i-f response characteristic, the stagger tuning method may be modified by tuning all or most of the coils to different frequencies. An example of this modification may be found in the video i-f amplifier shown in Figure 10. This i-f amplifier employs five stagger tuned circuits, each of which is tuned to a different frequency within the band. With the individual bandwidths and resonant frequencies properly chosen the desired over-all response is obtained.

The first tuned circuit consists of L_1 resonated with the output capacitance of mixer tube V_1 and input capacitance of the first i-f amplifier V_2. In a like manner, L_3, L_5, L_8 and L_{11} are resonated with the corresponding output and input capacitances in the 2nd, 3rd, 4th and 5th stages respectively. Insofar as the coupling action is concerned, it makes little difference whether the inductance element is located in the plate or grid circuit of the amplifier tube, and, in this amplifier, coil L_8 is located in the grid circuit of the last i-f tube and L_{11} in the cathode circuit of the video detector V_6.

Absorption traps are employed in the first four stages. In the first stage, trap L_2C_3 resonates at the accompanying sound intermediate frequency and, as indicated, a tap on L_2 is connected to the grid of the first sound i-f amplifier tube. Therefore, the sound and picture i-f's are separated in the mixer output circuit in a dual channel i-f system. For the remaining stages of the picture i-f amplifier, the over-all selectivity is adjusted to pass only the video i-f and its sidebands.

Trap L_4C_7 is tuned to absorb energy at the adjacent sound i-f. Trap L_6C_{12} is employed in a like manner to absorb energy at the adjacent video i-f.

Tuned to the accompanying sound i-f, traps L_7C_{18} and $L_{10}C_{19}$ prevent energy at this frequency from reaching the video detector. Absorbing energy at the accompanying sound i-f, $L_{10}C_{19}$ "reflects" resistance into the cathode circuit of V_5. This reflected resistance produces degeneration in the amplifier circuit so that the stage gain is greatly reduced at the sound intermediate frequency.

Combination Coupling Circuits

In some receivers, the video i-f amplifiers contain both bandpass filter and stagger tuned coupling systems as illustrated in Figure 11.

Here, between mixer tube V_1 and i-f amplifier V_2 and also between tubes V_4 and V_5, the coupling circuits are of the bandpass filter type, while stagger tuned circuits couple V_2 to V_3 and V_3 to V_4. Between V_5 and video detector V_6, a single tuned circuit consists of L_{11} and the circuit shunt capacitances. In the first stage, trap $L_3C_4C_5$ is tuned to reject the adjacent sound channel i-f, while in the fourth stage, trap $L_8C_{16}C_{17}$ rejects the accompanying sound i-f.

The action of rejector type traps is to offer high impedance at the frequency to be rejected. However, in the coil of the trap, I^2R losses tend to prevent complete blocking of the signal to be rejected. In the circuits of Figure 11, this fault is counteracted by resistors R_3 and R_{17} connected from ground to the center tap of the capacitance of the respective traps. These resistors produce a bridgelike action, causing the traps to be balanced at the respective rejection frequencies. When so balanced, the trap circuits have nearly infinite impedance, resulting in almost zero transmission of the respective rejection frequencies.

For removing the desired sound i-f from the video channel, tuned circuit $L_{12}C_{14}$ is connected so that a part of the coil is in series with R_{14} between the V_4 cathode and ground. The circuit is tuned to the accompanying sound i-f so that V_4 cathode current variations at this frequency will cause a large voltage to be developed across the coil.

From the ungrounded side of L_{12}, the sound i-f signals are coupled through C_{18} to the grid of the first sound i-f amplifier tube. With the sound i-f removed at this point, the circuits preceding this stage must have an over-all response broad enough to include both the sound and video sidebands, while those following need to pass only the video sidebands. Thus, this circuit is termed a modified dual channel system.

THE VIDEO DETECTOR

Operating on the same principle as the detector in an AM radio receiver, the video detector demodulates the video i-f to eliminate the intermediate frequency component. Its output contains only the video, blanking, and synchronizing signals at proportional amplitudes and exactly the same frequencies which they have at the input of the transmitter modulated stage.

Whether the video detector rectifies the positive or negative alternations of the applied i-f signal illustrated in Figure 12, depends upon the connections to the detector elements. This will determine the polarity of the resulting composite video signal in the detector output load.

Included in Figure 12 are the conventional series diode arrangements for television receiver detector circuits, along with the basic input and output waveforms. In many cases germanium crystals, such as the 1N34, 1N64, 1N87, etc. are used in place of a diode tube. They are indicated in dotted form in the diagram and can be inserted in place of the tube in the position specified.

Figure 12A shows the input plate circuit of a series video detector. If the wave-form of Figure 12C is applied to this input, with the diode V1 connected as shown, conduction occurs only during the positive alternations of the i-f input signal. Filtered by C2 and the output capacitance of the stage, the composite video signal resulting across R1 is illustrated at the right of the circuit shown.

It will be recalled that maximum amplitude of the carrier signal from the transmitter occurs during the sync-pulse period of time. In Figure 12C this is indicated as A. Maximum conduction of V1 will always occur when the i-f alternations reach this point.

Since diode conduction is from cathode to plate in Figure 12A, the pulsating diode electron path will be from the cathode to the plate, down through L_1 and back up through R1 to the cathode in a completely closed signal circuit as indicated by the arrows. Thus, during period A, maximum positive voltage will be developed across R1 by the i-f pulsations filtered by C2, to produce the positive sync pulse in the waveform at the right.

During any of the period of time indicated for the i-f alternations at the level shown by B in Figure 12C, approximately 25% less than maximum signal occurs resulting in a reduced filtered plus voltage across R1 to produce the blanking pulse as shown.

For that period of time indicated by C, the modulated i-f alternations representing the actual picture information are relatively lower in amplitude resulting in lower diode conduction levels and thus, lower positive voltages across R1 which, filtered by C2, produce the video variations between the blanking periods shown at the right in Figure 12A.

Since the picture information appears in the negative portion of the detected waveform shown, the polarity of this composite video signal is referred to as having negative picture phase, although the polarity across R1 always will be more or less positive for the entire composite video signal.

In Figure 12B, the diode is reversed so that the modulated i-f signal e_I is now applied to the cathode of the detector. Conduction of the diode now occurs on all the negative alternations of the input i-f signal.

When any a-c signal is applied to the cathode, it will cause conduction from cathode to plate on all the negative alternations. Point A′ in Figure 12C represents the maximum negative i-f alternations during the sync pulse period which will cause maximum diode conduction in an electron path as shown from cathode to plate in Figure 12B, down through the diode load R1′ and up through L_1', to the cathode. The detected and filtered i-f result during that time, will be a negative sync pulse, since a maximum negative voltage will result across R1′ with electrons going in the direction indicated.

During the blanking period level indicated by B′ the diode conduction for each alternation will be 25% less with a resulting less negative voltage across R1′ for the developed blanking pulse.

During the C′ period of time representing the picture information the amplitude of the i-f alternations in a negative direction are relatively lower resulting in lower diode conduction levels and thus low negative voltages across R1′ which, filtered by C2′, discharging through R1′, produce the video variations shown at the right in Figure 12B. Thus with the diode reversed, the pulsating d-c voltage across R1′ always will be more or less negative.

However, the picture information now appears in the positive portion of the detected waveform shown; therefore the polarity of this composite video signal is referred to as having positive picture phase.

Insofar as the video detector action is concerned, either circuit arrangement of Figure 12 is satisfactory. The number of video amplifiers inverting the signal between the detector and the picture tube along with the method in which the picture tube grid circuit elements are connected in the receiver, are all determining factors as to the way in which the video detector is to appear in its circuit. This relationship will be described more fully in the lesson on video amplifiers.

AUTOMATIC GAIN CONTROL

In many broadcast radio receivers, an avc circuit automatically reduces the total amplification of the signal when the strength of the received carrier wave increases, and increases the amplification when the carrier strength decreases. The change in amplification is accomplished by operating remote cut-off type r-f and i-f amplifier tubes with a grid bias voltage which is directly proportional to the average carrier amplitude. An increase in average carrier strength produces an increase in negative grid bias, thereby causing the amplifier tubes to operate on a less steep portion of their I_bE_c characteristics and produce less signal amplification. This amounts to varying the gain of the amplifier stages.

In the sound receiver, the avc is operating properly when it maintains a substantially constant average amplitude of signal input to the second detector. However, in the transmission of television signals, the carrier average amplitude charges in accordance with the average background brightness of the scene.

Thus, when a dark scene is transmitted, the average amplitude is high and when a bright picture is transmitted the average amplitude of the picture signal is low.

The sync-pulse peak, however, remains constant under all picture conditions and thus serves as an excellent reference for automatic gain control bias, referred to as agc. Therefore, the general method employed in the television receiver, is to develop a bias which is proportional to the peak carrier amplitude during the sync-pulse period, instead of the overall average amplitude of the video carrier.

If no fading or other effect causes a variation in peak carrier strength, this bias and, therefore, the gain of the amplifiers remains constant. However, any peak carrier amplitude change will result in a compensating change of agc bias and signal amplification.

AGC Circuits

Figure 13 shows the diagram of an automatic gain control circuit. V_1 is the last i-f amplifier tube, V_{2A} is the agc detector and V_{2B} is the video detector. The i-f signal appears across L and is applied to V_{2B} cathode for the usual video detector action as described for Figure 12B. The i-f is coupled also through C_1 to the plate of an agc diode.

The positive i-f signal alternations, during period A, cause the diode V_{2A} to conduct, producing electron movement in the direction indicated by the black arrows. This charges the agc capacitor C_1 negatively as indicated by the polarity shown.

During the intervals between the i-f signal positive peaks of which there may be several hundred during one horizontal sync pulse period, the diode becomes non-conductive with C_1 discharging only very slightly through R_1 and the coil L. However, at the end of each sync-pulse period the i-f amplitude is less for the blanking and picture information periods as shown by the wave-form in Figure 13; now capacitor C_1 will discharge in the direction indicated by the dotted arrows, until the next sync pulse period arrives.

Because C_1 and R_1 in Figure 13 are comparatively large, the discharge circuit time constant is relatively long and C_1 discharges slowly for the entire line of picture information between sync pulses to maintain a high negative voltage across R_1 as indicated by the polarity markings. This action is identical to the grid-leak bias action described in earlier lessons.

During this time between sync-pulse periods the negative voltage maintained on the diode plate across R_1 is greater than the positive alternations of the i-f signal. Thus, whether the signal represents bright or dark picture information the diode remains cut-off until the next sync pulse period.

The maximum and minimum voltages across R_1 are indicated in the wave-form in Figure 13 as E_{R_1}, which is actually a pulsating d-c with a frequency of approximately 15,750 cps. This negative voltage is filtered by R_2C_2 of Figure 13 whose output, then, represents a voltage proportional to carrier peak power to be used as the automatic gain control bias for the i-f and r-f stages where it is applied to their grid return circuits.

Thus, an increase in the i-f amplifier output peak causes greater conduction of the agc diode and an increase of negative agc bias voltage. This greater negative bias reduces the amplification of the r-f, i-f stages to bring the signal level back to normal. In a similar manner, a decrease in the signal peak during the sync periods causes less conduction of the diode for a decrease in negative potential and an increase in the amplification of the r-f and i-f stages to bring the signal level back up to normal.

Occasionally, R_1 may be two resistors in the form of a divider to ground for the purpose of obtaining a lower negative control voltage for the r-f tuner, so that the gain and thus the signal-to-noise ratio remains relatively high in the r-f stages. In that case another filter such as R_2C_2 would be added from that divider point.

In Figure 14 is a diagram including an agc circuit whose output is varied by contrast control action. V_1 is the last i-f amplifier tube, V_{2A} is the agc diode, V_{2B} the video detector, and V_3 is the v-f output amplifier. The V_1 output appears across L and is coupled through C_2 to the plate of the agc diode section of V_2.

The positive i-f signal alternations cause the diode V_{2A} to conduct, to charge the agc capacitor C_2 as the polarity signs indicate.

In parallel with C_5, series connected resistor R_4 and a part of potentiometer P are also in the electron path. Because the diode and coil L have very low resistances and C_5 is large in comparison to C_2, C_2 charges very rapidly to the polarity indicated and very nearly equal to the i-f signal peak. During the intervals between the sync-pulse periods, the diode becomes nonconductive and C_2 discharges through R_3 and the coil L. The resistance of R_3 is again very large so that the discharge circuit time constant is relatively long and the capacitor discharges slowly to maintain a high negative voltage across R_3.

In the circuit of Figure 14, a 1K potentiometer P serves as the video amplitude or contrast control, and can change the magnitude of the developed agc bias from V_{2A}, and the amount of degeneration in the cathode circuit of video frequency output tube

V_3. Resistor R_5 and the part of P between it and the slider form a voltage divider between B+ and ground as indicated. Connected through R_4 to the junction between R_5 and P, the cathode circuit of the agc diode therefore has a d-c bias which is positive with respect to ground.

Thus, the agc detector V_{2A} stays cut off until the positive i-f signal peaks on the plate are greater than the plus voltage on the cathode. The extent that C_2 charges through V_{2A} for the proportional agc bias depends on how much more positive the i-f peaks on the plate are compared to the positive bias on the cathode.

When the slider is moved toward R_4, this cathode positive bias is reduced and the conduction of the agc diode is greater during sync periods. This action causes a higher agc voltage which decreases the amplification in the r-f and i-f stages. This same adjustment of the slider results in a greater resistance in the V_3 cathode circuit, increasing the amount of degeneration, thereby reducing the over-all amplification of the signal some more. The reverse is true also; moving the contrast control slider toward R_6 causes an increase in amplification by the r-f, i-f, and v-f output stages.

The contrast control, producing similar results to the picture as the volume control does to sound, often is incorporated into a bias circuit that affects the gain of the r-f and i-f stages.

The purpose of the C_5-R_4 combination in the cathode circuit of V_{2A} is to assist in maintaining diode cutoff between sync periods. C_5 charges to the polarity indicated when V_{2A} conducts, and discharges through part of control P and R_4 during the blanking and video time for the self bias action. This will reduce the tendency of V_{2A} to conduct for noise pulses.

Amplified AGC

Some receivers incorporate an agc voltage amplifier to increase the sensitivity of the system and, as an example of this arrangement, Figure 15 shows another agc circuit. Here, V_1 is the last i-f amplifier tube, V_2 is the video detector and agc detector, V_3 is the agc amplifier and V_4 is the agc diode.

In order to produce the desired bias, this entire circuit is operated below ground potential and is supplied with negative d-c operating voltages from the power supply. In operation, the i-f signal that is developed across coil L is coupled through C_2 to the agc detector, where the positive signal alternations cause the tube to conduct and charge C_3 to a value E_{C_3} which is nearly equal to the i-f signal peak.

The agc detector output, E_{C_3} is applied through the long time constant filter $R_3R_4C_4$ to the grid of V_3 which operates as a direct coupled amplifier. As shown, the V_3 plate connects through R_{10} and R_{11} to ground. However, normal operation of this tube is obtained since its cathode is connected to the slider on potentiometer P, and, as indicated, it is 99 volts negative with respect to ground. Likewise, returned through R_4, R_3 and R_2 to the -105 volt power supply terminal, the V_3 grid is operated 6 volts negative with respect to the cathode, less the voltage E_{C_3} produced by the received signal.

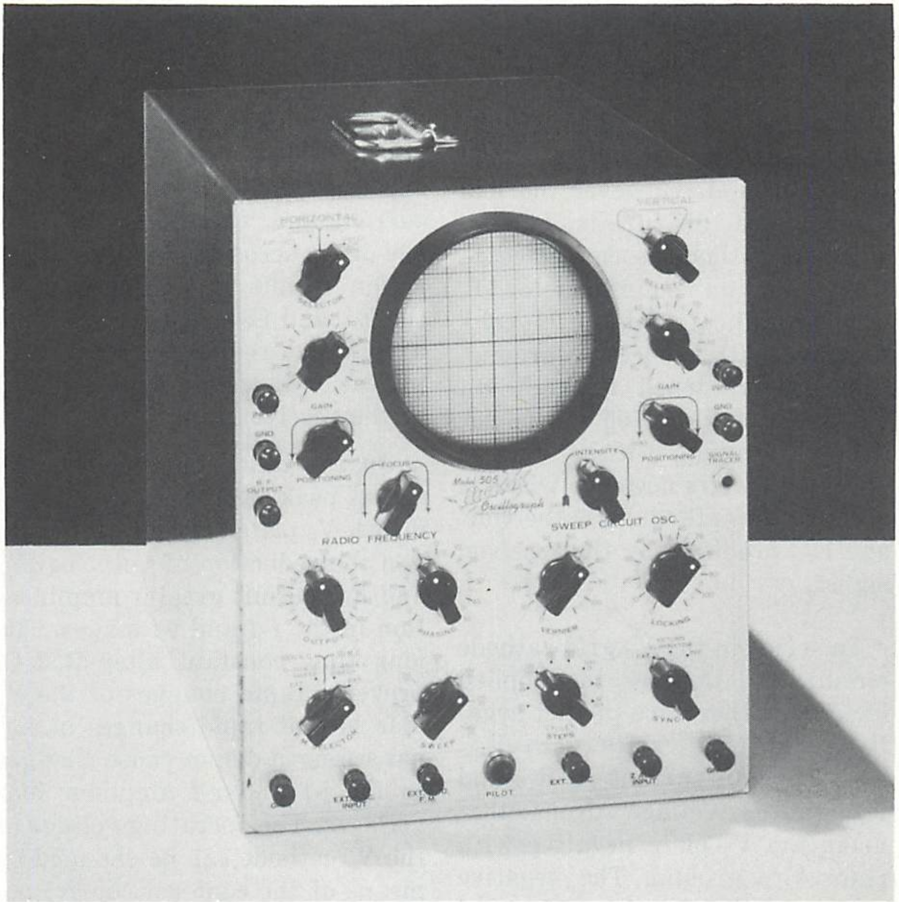
In series in the V_3 grid-cathode circuit, E_{C_3} opposes the applied fixed bias, therefore the net negative grid bias varies with E_{C_3}. The V_3 plate current in R_{10} and R_{11} causes voltage drops that make the V_3 plate negative with respect to ground. The negative voltage at the junction of R_{10} and R_{11} is filtered by C_7, which charges to the polarity shown, and is applied to the grid return of the r-f amplifier tube.

Capacitor C_6 is connected across both resistors and charges to their total voltage drop with the polarity shown. The voltage across C_6 is applied through filter R_9C_5 to the grid returns of the i-f amplifier tubes.

In operation, an increase in the i-f signal peak causes greater conduction of the agc detector during the sync pulse intervals, charging C_3 to a higher voltage. The increase of E_{C_3} makes the V_3 grid less negative, increasing the plate current and, therefore, the voltage drops across R_{10} and R_{11}. This action results in an increase of the E_{C_6} and E_{C_7} negative agc voltages, which reduce the signal amplification in the r-f and i-f stages of the receiver.

On the other hand, when the signal peak decreases, E_{C_3} is reduced, V_3 plate current decreases, and the reduction of voltages E_{C_6} and E_{C_7} permit greater amplification in the r-f and i-f stages. The long time constant filter $R_3R_4C_4$ prevents rapid changes of the V_3 bias so that rapid changes in signal strength do not cause changes in the i-f and r-f amplifier bias voltages. The operating voltage of the V_3 cathode can be changed by means of the contrast control potentiometer P_1, thereby permitting manual adjustment of the agc voltages developed. This eventually controls the amplitude of the signal at the picture tube.

Since capacitor C_6 is connected across both R_{10} and R_{11}, a higher voltage is developed across it than across C_7. For very high input signal amplitudes, the agc bias applied to the i-f amplifier grids may be so high as to cause the



The actual response curve of video i-f amplifiers can be reproduced on an oscilloscope screen when used with a sweep generator.

Courtesy Hickock Electrical Instrument Co.

tubes to operate near the cut-off point and distort the signal. To prevent this distortion, agc clamping diode V_4 is employed to limit the bias applied to the i-f amplifier tubes by connecting its cathode to the i-f agc line and its plate to a −7.2 volt power supply terminal.

Normally, the V_3 plate and therefore the V_4 cathode is less negative than −7.2 volts, and V_4 is nonconductive. However, when E_{c_6} exceeds −7.2 volts, the V_4 cathode becomes more negative than the plate and the diode conducts. When conductive, a diode has relatively low resistance and resistor R_9 and the V_4 plate resistance form a voltage divider from the negative plate of C_6 to the -7.2 volt terminal. The i-f agc line is connected to the junction between V_4 and R_9, therefore, when V_4 is conducting, the voltage applied to the i-f amplifier grids will be only a fraction of the total voltage across C_6. That is, after E_{C_6} reaches -7.2 volts, there is a considerable reduction in the rate of agc voltage rise.

The V_3 duo-diode plate circuits prevent changes in agc voltage due to high amplitude transient voltages arising from automobile ignition "noise," etc. Such voltages cause sudden heavy conduction of the agc detector, and the voltage drop E_{R_2} becomes sufficiently high to make the diode plates of V_3 positive with respect to the cathode. The resulting electron flow, from the cathode to the diode plates, cause voltage drops across R_2, R_3, R_4, R_5 and P, all of which are of a polarity which tends to make the grid negative with respect to the cathode and prevent a rise of V_3 triode plate current.

The agc circuits just described are fairly satisfactory when all stations can be received with minimum noise and when changes in signal strength occur slowly. The filter networks that smooth out the agc voltage must have RC time constants sufficiently long to remove the 60 cycles per second vertical synchronizing pulses. Therefore, when the changes in signal strength occur in about 1/60 of a second, or faster, the bias voltage does not change at all.

A serious drawback exists in these basic agc circuits when the noise level is very high. The noise itself causes diode conduction between sync pulse periods producing a more negative bias and thus reducing the gain of the controlled stages. In areas where a weak signal station is received with a high level of noise, there is a tendency to suppress the television signal altogether, since the noise can produce a bias so large that the already weak signal does not receive sufficient amplification.

In an effort to overcome these drawbacks, a fast acting agc circuit has been developed, known as keyed agc. Its main objective is allowing the agc circuit to develop bias only during the sync pulse periods. Its keying circuit prevents high amplitude noise pulses, that occur during the blanking and video time, from producing a false negative bias.

Since this circuit depends on other receiver circuits yet to be described, it is described in a later lesson.

A modification of these agc circuits for television is shown in the circuit in dotted form connected out of the video detector in

Figure 13. Since the negative alternations of the i-f signal cause V_{2B} to conduct, a pulsating d-c, filtered by C_3, will appear across the detector load R_3 with the negative polarity indicated and the positive picture phase shown.

The resulting negative d-c voltage varying at the composite signal rate is then filtered by R_4-C_4 for bias action. However, this agc bias is an average condition which will change with picture brightness, as mentioned at the beginning of this subject on agc. It is not a perfectly desirable situation but it does provide a means for general automatic bias reference that reduces with a weak over-all signal and increases with a strong one. This method of agc, requiring only a resistor and capacitor, is inexpensive and therefore used in quite a number of cases to help keep down the price of the receiver.

A good sweep signal generator is essential for accurate reproduction of the i-f amplifier response curve on an oscilloscope screen.
Courtesy RCA Victor

AGC and I-F Circuit Defects

If, for any reason, the agc circuit becomes defective so as to reduce negative bias in the i-f stages to practically zero, various types of distortion of the composite signal can occur, depending upon the specific signal circuit arrangement.

For example, in Figure 8, if the agc filter capacitor C_{11} shorted, the bias on stage V_3 would be greatly reduced, while that on V_2 would be completely reduced to zero. The gain of these stages would be maximum which would result in a very high amplitude composite video signal at the plate of the video detector V_5.

V_5 can be compared to V_{2B} in Figure 14 which shows the complete video detector circuit. The very large i-f signal applied to the cathode of the video detector will cause very heavy diode conduction and thus produce a large average negative d-c voltage across the diode load R_7 as shown.

This negative d-c voltage can be great enough when directly coupled to the video amplifier to place the grid circuit at cutoff, for no picture signal output at the plate of V_3 and thus, a no-picture condition on the screen of the picture tube. If the receiver is of the dual channel type, the sound will not be appreciably affected, whereas with the intercarrier type, the point of take-off for the 4.5 mc sound i-f signal will determine the extent to which the sound is affected.

If another channel with a lower signal level is selected by the tuner, the negative voltage across the diode load R_7 will be less and may allow some of the peaks of the video information to appear in the plate circuit of V_3; however, the sync and blanking pulses would still be clipped out; thus the picture information could not be synchronized and would appear broken up on the picture tube screen.

In a number of these cases where the agc bias has been lost, if the antenna is partially or completely disconnected, a picture will then appear on the picture tube. This result occurs because the signal input to the receiver, picked up by part of the transmission line, is now very low and does not become great enough in amplitude at the output of the video detector to develop signal cut-off bias for V_3. Noise, of course, will be maximum so that this resulting picture will have much noise disturbance in it and be, as termed, "snowy" in appearance.

In circuits where capacitive coupling is used from the video detector instead of direct coupling, the extremely high peak to peak amplitude of the composite signal resulting from zero agc bias may overdrive the video amplifier into saturation and cutoff to distort the output signal. The sync pulses will be clipped to cause poor synchronization and the video information will have maximum amplitude with its peaks clipped to produce a picture, if it can be synchronized, of poor detail with heavy contrast of black and white.

One of the common causes of agc bias failure, is a defective i-f tube. If for some reason the control grid should begin emitting electrons, the agc circuit will loose electrons to that grid emission point. Thus the bias for all the r-f and i-f stages controlled by agc will be reduced close to zero to produce the undesirable effects just mentioned.

As mentioned previously, various kinds of distortion can occur from agc troubles depending upon circuit arrangements in the particular receiver. For example, in Figure 14, if V_{2A} became nonconductive or the 100 $\mu\mu$f capacitor C_2 which couples the i-f to the plate circuit of the agc detector opened, again there would be no agc bias output from the circuit. The extremely maximum signal now occurring in the grid circuit of the last i-f stage would cause a large grid-leak bias across R_9.

This bias could be so great as to almost completely cut off the i-f tube allowing only scatterings of unintelligible picture information to appear on the picture tube screen.

On a weaker channel signal, it is possible that a negative picture would result with this agc defect. That is, all the dark portions of the signal would be white and all the white portions dark. This inversion is the result of the high grid leak bias action in the grid circuit of the last i-f stage.

Since maximum grid-leak bias occurs on peak signal which is during the blanking and sync pulse period, the gain of the stage will be minimum for this period of time. During the time of picture information, the i-f signal representing the dark portions of the signal will develop high grid-leak bias for minimum gain of the stage while the low i-f signal representing the whites in the picture information will produce the least grid-leak bias to allow the i-f to have maximum gain. In the output of the last i-f stage, this results in an inversion of the modulated wave-form, as sketched in Figure 16A. It should never appear anywhere in any stage of a television system.

When this incorrect signal is applied to the video detector cathode in Figure 14, the detected result, illustrated in Figure 16B, is inverted from the proper wave-form shown in Figure 14 and will cause the negative picture condition. These, then, are some of the varied results which can occur from a defective agc circuit.

If a screen grid by-pass capacitor should open in one of the last i-f stages, there is sometimes a possibility that a similar inversion of the signal as described for agc defects, will occur due to screen grid detection. This would produce maximum degeneration during the peak of the i-f signal thus causing reduced stage gain, while it would be inverse for the low portions of the i-f signal.

Another result of an open screen capacitor would be poor response of the dark portions of the picture information since the i-f signal is maximum during that time causing maximum degeneration. In addition, the resulting picture information would be poor in detail because of the poor coupling, stray or otherwise, of a small portion of this unwanted detected signal into the video detector and amplifier circuits. This can cause a smearing condition in the picture.

However, in most cases, when a screen and plate circuit capacitor opens, the i-f circuit oscillates or rings, and produces a succession of multiple images in the resulting picture, which look like shadows appearing to the right of the original image.

Other troubles such as the complete interruption of i-f stage operation resulting from a defective tube or other component may cause either a zero signal at the output of the receiver, or sometimes a low signal output producing a very weak picture. When any signal gets through a defective stage, it is the result of interelectrode and stray capacitive coupling of the carrier signal from the stage preceding the defective one to the stage following it. This condition could occur for any of the r-f or i-f stages.

If the receiver is of the intercarrier type, then the sound output also is affected. However, in the dual or modified-dual channel receiver, if the sound signal has been taken off ahead of the defective stage, it remains normal.

The video amplifier, described in the next lesson, is the last link in the series of picture signal amplifying sections for the picture information to the picture tube. It is to the picture tube of a television receiver what the audio section is to the speaker. However, where the frequency response of the audio section may be from a few cycles to 10 kc or 15 kc, that of the video amplifier section must extend from approximately zero to 4,500 kc or 4.5 mc.

This frequency range requires careful circuit considerations because the ratio of frequency coverage between maximum and minimum is so extreme. In addition the polarity of the composite video signal takes on an important aspect.

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STUDENT NOTES

STUDENT NOTES

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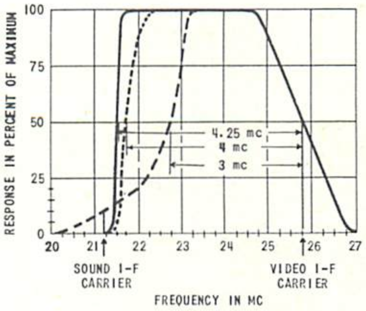


FIGURE 1

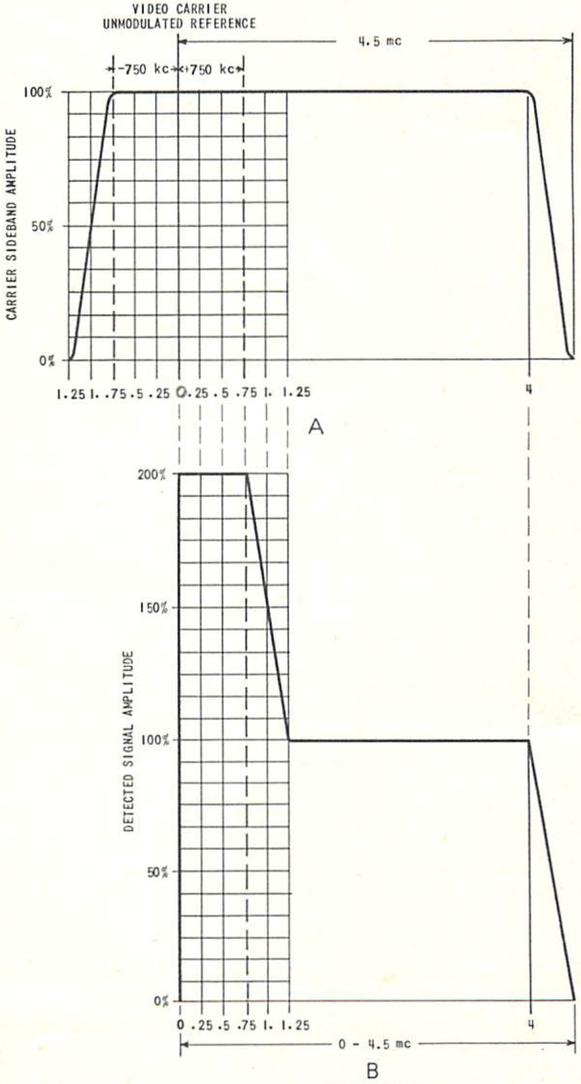


FIGURE 2

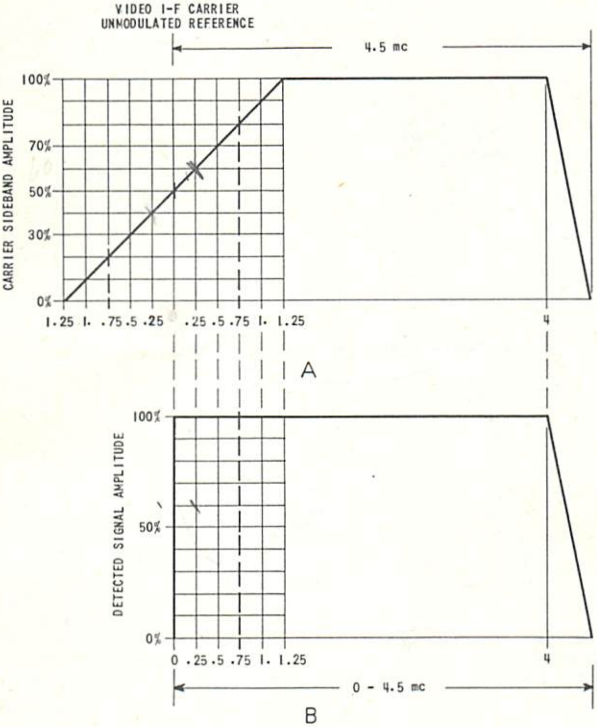


FIGURE 3

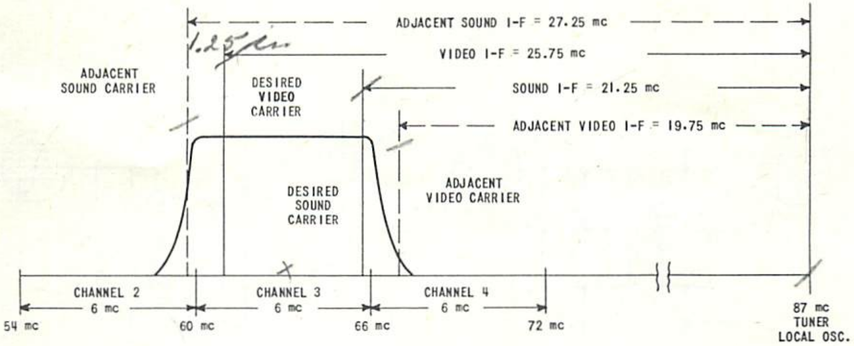


FIGURE 4

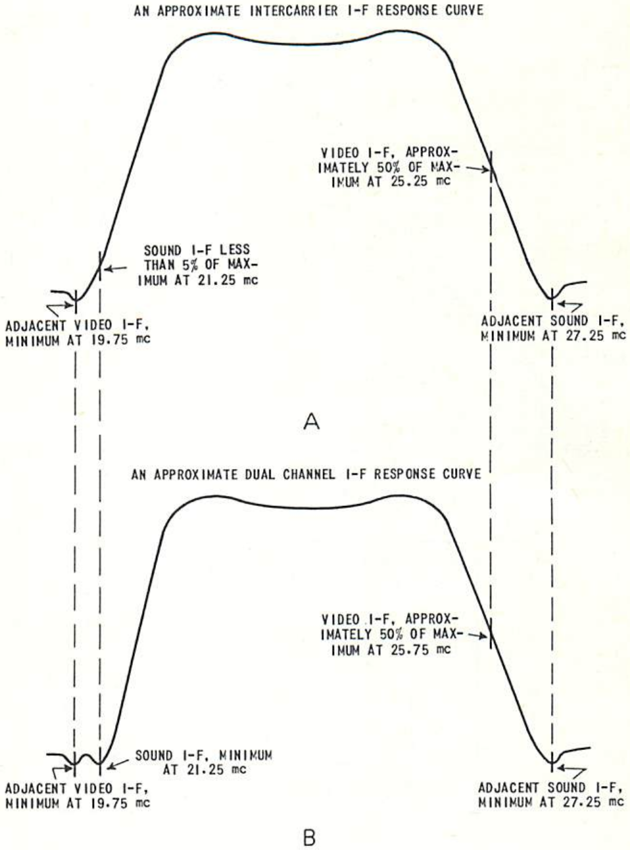


FIGURE 5

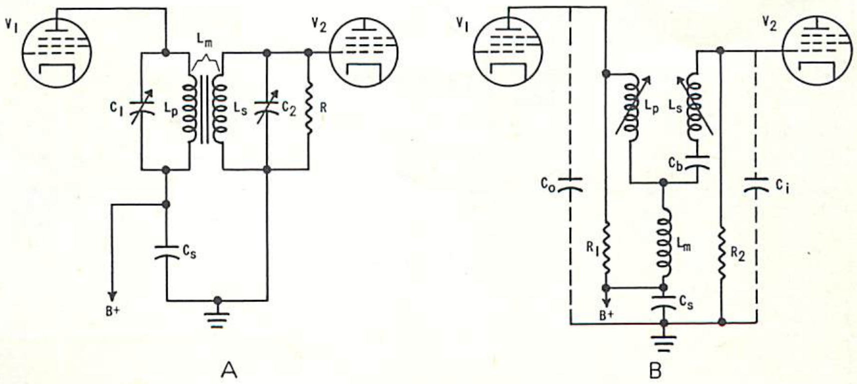


FIGURE 6

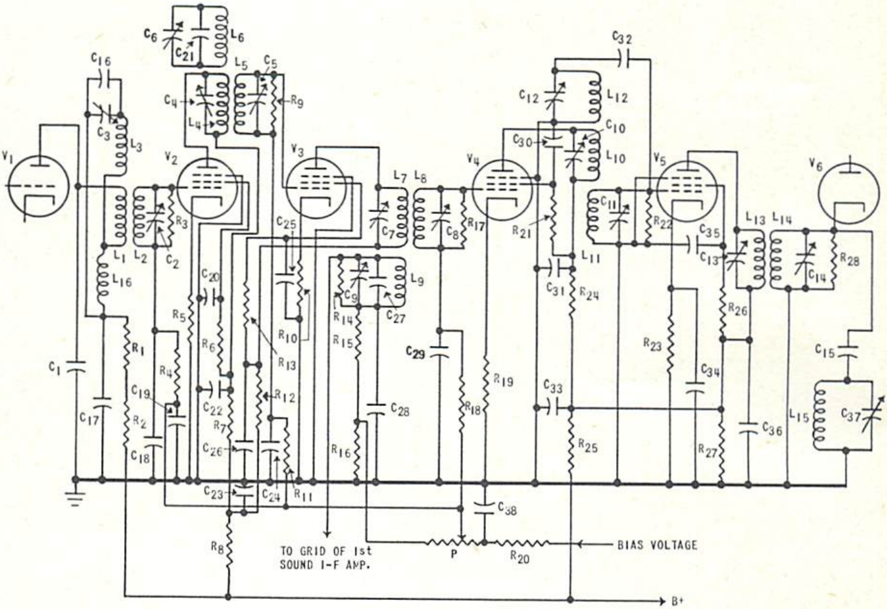


FIGURE 7

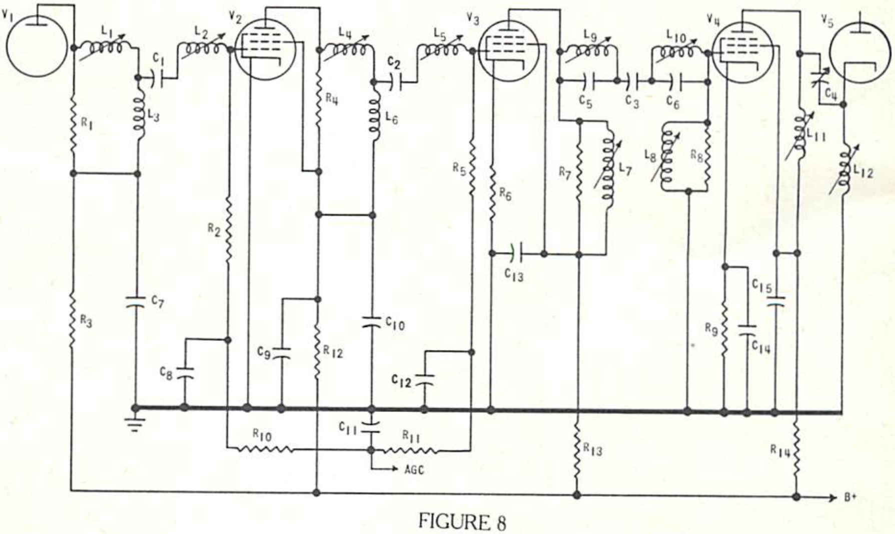


FIGURE 8

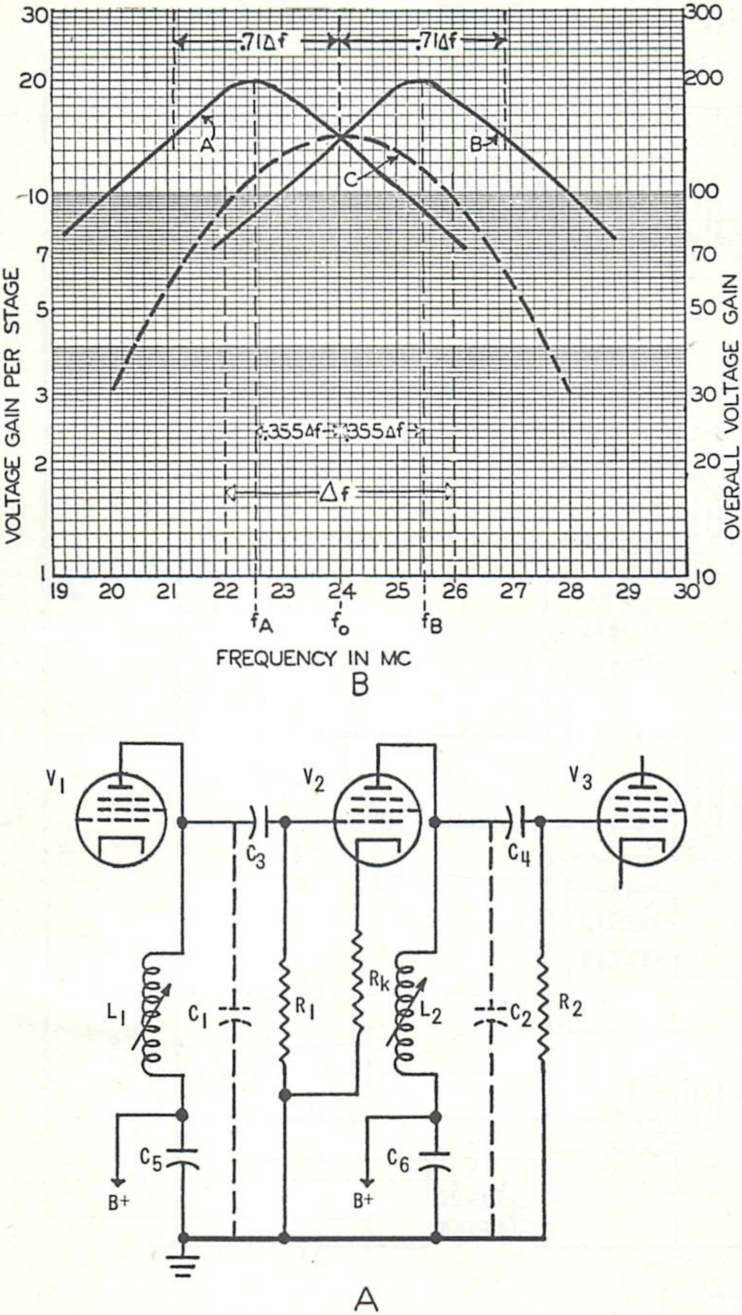

30
20
10
7
5
3
2
1
300
200
100
70
50
30
20
10
VOLTAGE GAIN PER STAGE
OVERALL VOLTAGE GAIN
.71Δf
.71Δf
A
B
C
.355Δf
.355Δf
Δf
19
20
21
22
23
24
25
26
27
28
29
30
f_A
f_o
f_B
FREQUENCY IN MC
B
V_1
V_2
V_3
C_3
C_4
L_1
C_1
R_1
R_K
L_2
C_2
R_2
C_5
C_6
B+
B+
A

FIGURE 9

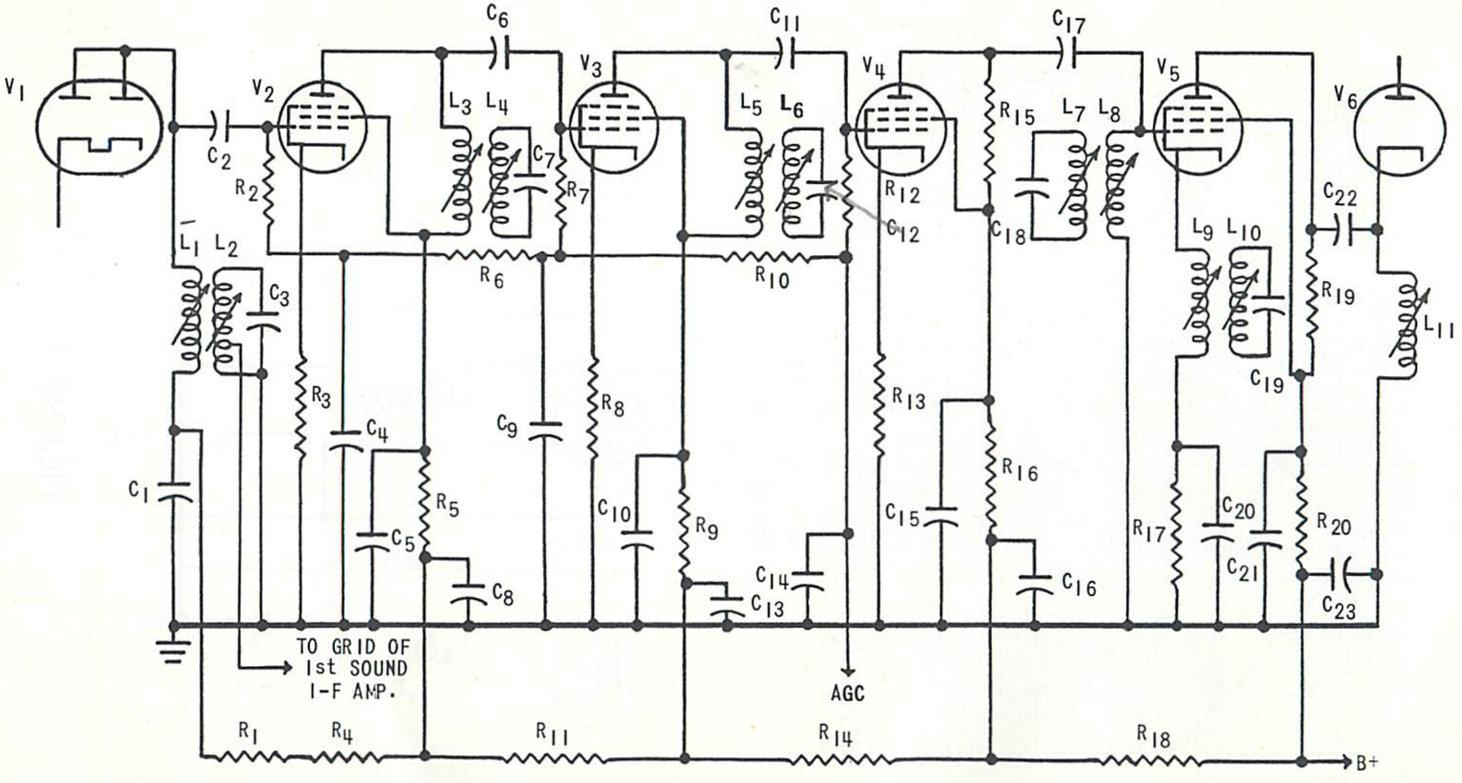

V1
V2
V3
V4
V5
V6
C1
C2
C3
C4
C5
C6
C7
C8
C9
C10
C11
C12
C13
C14
C15
C16
C17
C18
C19
C20
C21
C22
C23
L1
L2
L3
L4
L5
L6
L7
L8
L9
L10
L11
R1
R2
R3
R4
R5
R6
R7
R8
R9
R10
R11
R12
R13
R14
R15
R16
R17
R18
R19
R20
TO GRID OF
1st SOUND
I-F AMP.
AGC
B+

FIGURE 10

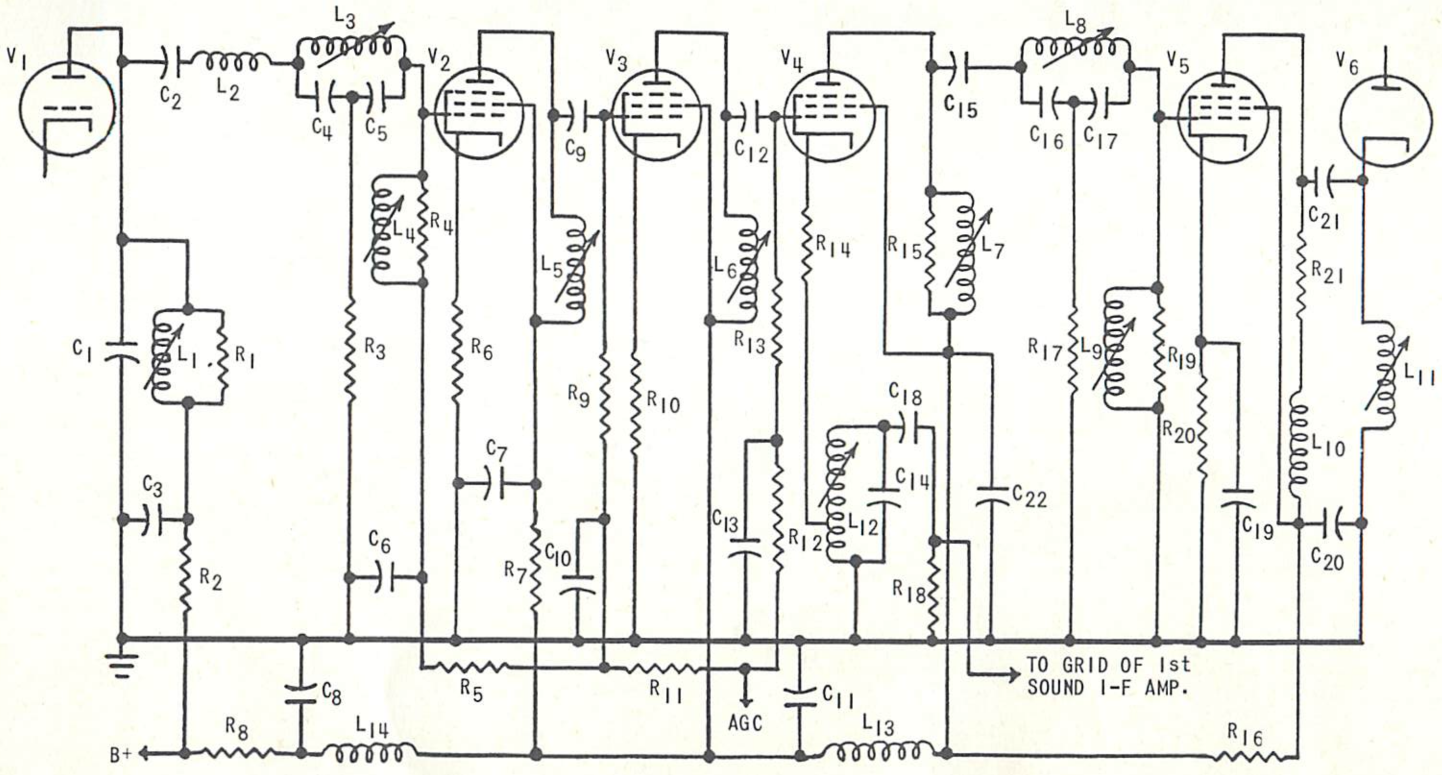

V$_1$
V$_2$
V$_3$
V$_4$
V$_5$
V$_6$
C$_2$
L$_2$
L$_3$
C$_4$
C$_5$
L$_8$
C$_{15}$
C$_{16}$
C$_{17}$
C$_9$
C$_{12}$
C$_1$
L$_1$
R$_1$
L$_4$
R$_4$
R$_3$
R$_6$
L$_5$
R$_9$
R$_{10}$
L$_6$
R$_{13}$
R$_{14}$
R$_{15}$
L$_7$
R$_{17}$
L$_9$
R$_{19}$
R$_{20}$
C$_{21}$
R$_{21}$
L$_{11}$
L$_{10}$
C$_{18}$
C$_{14}$
C$_{22}$
C$_{19}$
C$_{20}$
C$_3$
C$_7$
C$_6$
R$_2$
R$_7$
C$_{10}$
C$_{13}$
R$_{12}$
L$_{12}$
R$_{18}$
R$_5$
R$_{11}$
C$_8$
C$_{11}$
AGC
TO GRID OF 1st
SOUND I-F AMP.
B+
R$_8$
L$_{14}$
L$_{13}$
R$_{16}$

FIGURE 11

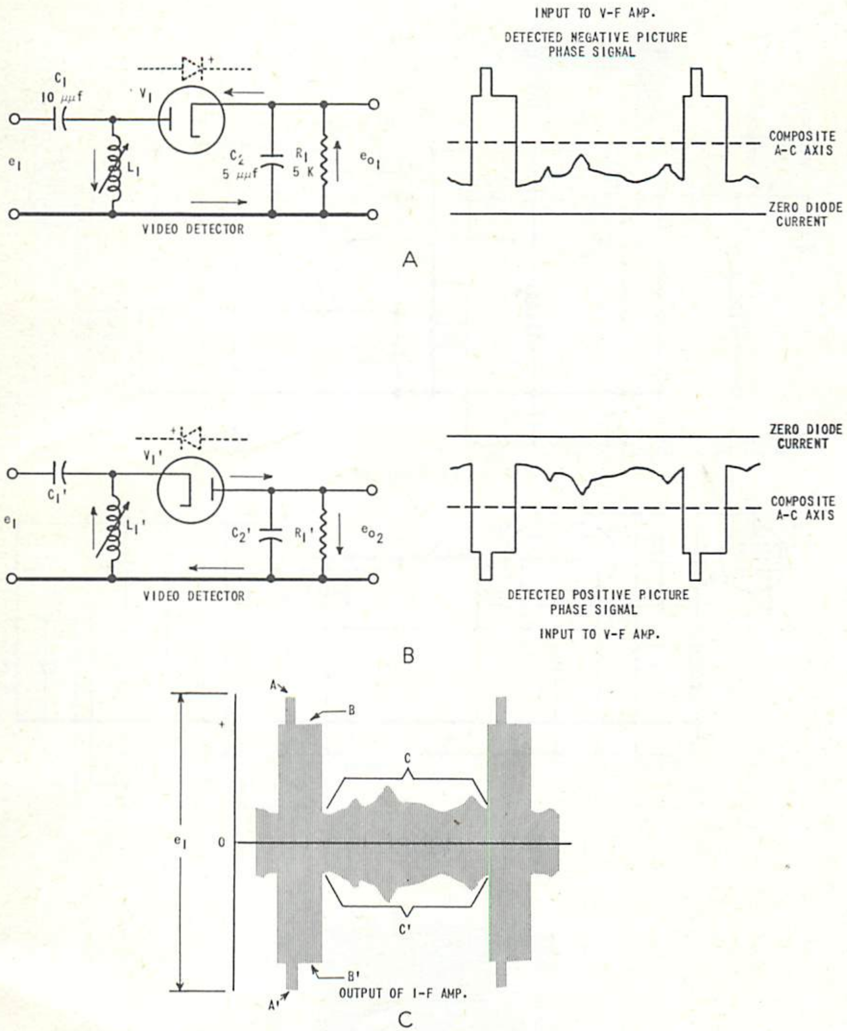


FIGURE 12

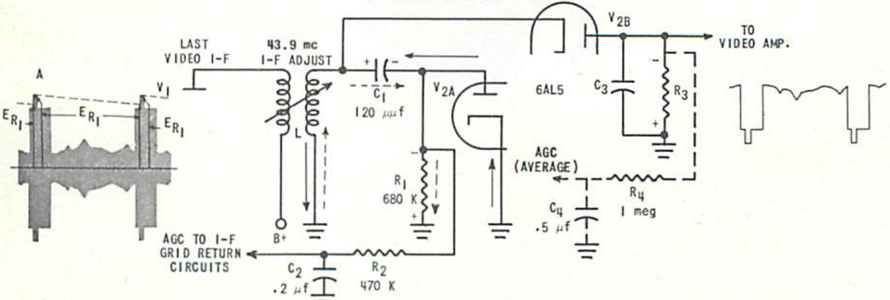


FIGURE 13

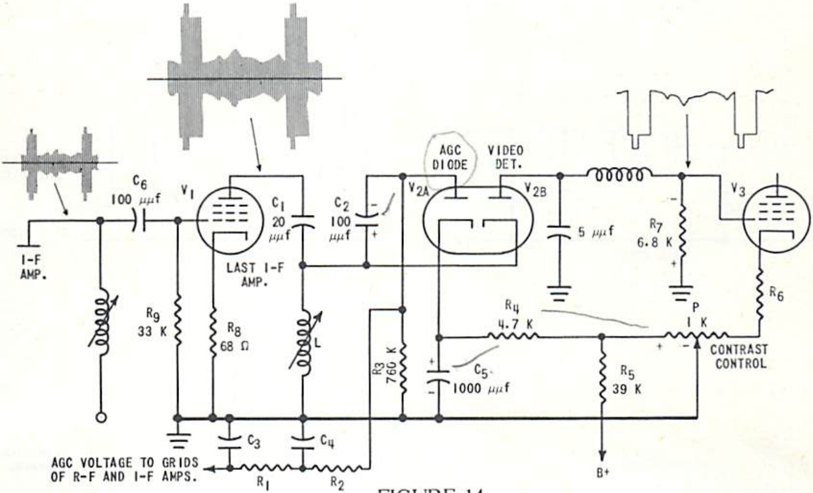


FIGURE 14

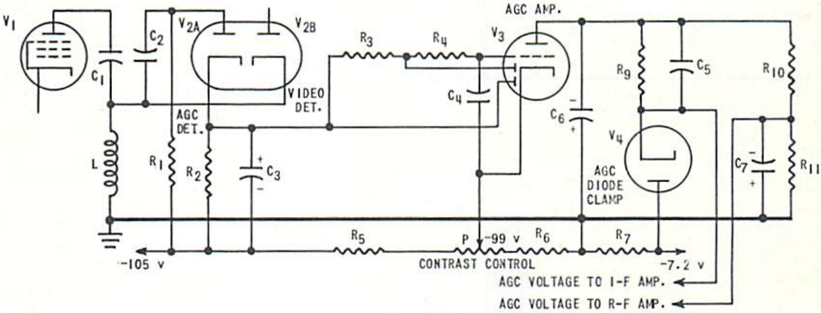
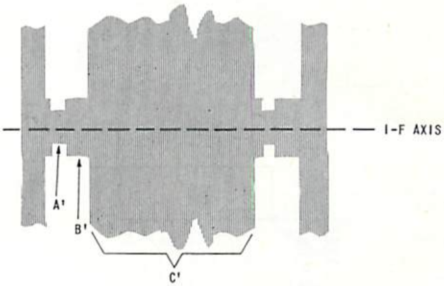
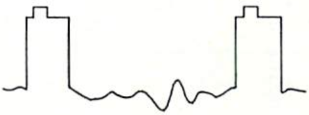


FIGURE 15



AGC DEFECT CAUSING I-F INVERSION

A



B

FIGURE 16

FROM OUR *Director's* NOTEBOOK

SUCCESS COMMANDMENTS

1. **WORK HARD.** Hard work is the best investment a man can make.
2. **STUDY HARD.** Knowledge enables a man to work more intelligently and effectively.
3. **HAVE INITIATIVE.** Ruts often deepen into graves.
4. **LOVE YOUR WORK.** Then you will find pleasure in mastering it.
5. **BE EXACT.** Slipshod methods bring only slipshod results.
6. **HAVE THE AMERICAN SPIRIT OF CONQUEST.** Then you can successfully battle with and overcome difficulties.
7. **CULTIVATE PERSONALITY.** Personality is to a man what perfume is to a flower.
8. **HELP AND SHARE WITH OTHERS.** The real test of business greatness lies in giving opportunity to others.
9. **BE DEMOCRATIC.** Unless you FEEL right toward your fellow men, you can never be a successful leader of men.
10. **IN ALL THINGS DO YOUR BEST.** The man who has done his best has done everything. The man who has done less than his best has done nothing.

—Charles M. Schwab

Yours for success,

W. C. DeVry

DIRECTOR

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